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A Spectrum Analyser for the Radio Amateur

Spectrum analysers enjoy a special place in the hearts and minds of amateur radio constructors. For one thing, they are able to display immediately the full output spectrum of a transmitter and the relative amplitudes. The other awe provoking thought which springs to mind is their enormous cost – now, almost as much as a small house. The possession of a spectrum analyser must surely be, for most amateur constructors, completely out of the question. Also, a perusal of a commercial analyser's specifications are enough to convince many that the amateur construction of such an instrument is fraught with unsurmountable difficulties. But, of course, one should'nt despair quite so easily because, for example, what radio amateur requires facilities such as a 10 Hz resolution at 20 GHz? In other words, if the facilities offered by a commercial spectrum analyser were pruned down to those required by the radio amateur then it is indeed entirely feasible to undertake the home construction of such an instrument.

The following article is intended to excite the reader to experiment with the project, those who require a "watertight" construction article complete with PCB layout patterns and a guarantee of sure-fire results will be disappointed. It must be appreciated that the detailed description of such a project would occupy all the pages of VHF COMMUNICATIONS for the whole year. The author is, however, prepared to give advice to various project-groups which might be formed.

In the almost continuously occupied frequency bands of today, a smooth succession of stations carrying various services can only be maximized within a given band if these stations observe the minimum demands concerning the radiation of extraneous energy. Even modern high-level receivers employing ring mixers fed with a high oscillator power in the region of 200 mW can, when poorly designed/constructed, cause a lot of trouble. Whilst one can be reasonably assured that a transceiver of a proprietry manufacturer will satisfy at least the minimum requirements of the radio regulatory authority, it will not be so certain that a home-constructed piece of equipment will offer the same freedom from spurious and unauthorized radiation. This can only be ascertained, in most cases, by a visit to the post office stand at a large ham-fest where the item may be subjected to the statutory tests.

If a complete survey of the harmonic and intermodulation content of a transmitter or oscillator is required then the spectrum analyser is the correct instrument to do it. It represents an

electronically periodically tunable superheterodyne receiver which is able to display the level of the signal-under-test together with the relative levels of its modulation and spurious signals. A commercial instrument would cover, typically, a range from a few Hz right up to twenty GHz (with supplementary mixers 325 GHz) and cost from 20 to 300 thousand DM. Not many amateurs could afford to buy a new one and those offered for sale at flea-markets are ageing, obselete examples which have a restricted dynamic range. Those not having the luck to find a good second-hand analyser might consider the idea of making one for themselves.

In 1976, DL 8 ZX published one of the first home-brew concepts which covered a frequency range from 0 to 60 MHz and from 120 to 180 MHz. A cable TV tuner was introduced in 1980 which was used as the first down-converter. The present article will consider how a spectrum analyser with usable specifications can be realized with a tenable degree of constructional complexity.

1. CONCEPT

Before a circuit concept may be considered, it is necessary to define the facilities the instrument should offer in order that a sense of proportion is acquired and that no subliminal objectives should be striven for. A few minimal demands will therefore be set down as follows: –

Frequency range: 0 – 500 MHz and eventually 500 – 1500 MHz. This covers all the chief activities up to 70 cm with the basic unit. The additional down-converter covers the 23 cm band and all the important signal processing frequencies for the production of signals in the microwave bands.

Dynamic range: At least 60 dB. A harmonic and intermodulation capability of 60 dB is perfectly sufficient and is, indeed, better than that attained by some lower-priced commercial instruments. Valved power amplifiers, without output filtering, achieve only up to 40 dB harmonic and intermodulation specification and 2 m and 70 cm transistor PAs are not much better.

Resolution (analyser bandwidth): Switchable from 200 – 500 kHz for survey measurements and down to 1 kHz in order to identify third-order intermod. products and neighbouring synthesizer channels.

Stability: Short-term stability must be better than the smallest resolution bandwidth i.e. 1 kHz. Long-term stability: better than ± 3 dB both over the whole frequency and the whole dynamic range.

Sensitivity (10 dB s = N/N): better than – 100 dBm i.e. 12 μV/50 Ω at the smallest bandwidth.

LO spectral purity: Noise sidebands ± 25 kHz from carrier should be lower than – 60 dB in order to preserve the validity of the dynamic range specification.

The circuit should use easily obtainable components and require the minimum of tuning adjustment even if this might mean increasing the circuit complexity somewhat. Otherwise, the circuit might require the services of another spectrum analyser to align it – and that doesn't help anyone. A few modules will now be regarded a little more closely in order to evolve the various realization possibilities and at the same time identify a few potential trouble-spots in order to circumvent them. The most important step in this direction is the determination of the local oscillator and IF frequency plan.

1.1. Frequency Plan

The spectrum analyser is a periodically, tunable receiver with an extremely large frequency range encompassing two or three decades. Signals within this frequency range must be converted into a fixed intermediate frequency in order that the selection and amplitude processing can be carried out. Every superhet. receiver has a principal spurious receive frequency, known as the image-frequency whose effects are normally rendered harmless, in a conventional receiver, by the preselector filter circuits. It goes without saying that this technique is not suitable for this application where the tuning range is completely continuous. The only solution therefore is to

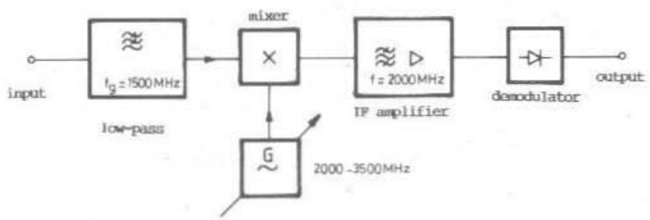

Fig. 1.1: Block diagram of 0 - 1500 MHz spectrum analyser

make the IF, as with many modern general-purpose receivers, lie above the highest receiver input frequency. The image frequency can then be identified and filtered out of the receiver input by means of a simple low-pass filter. The image frequency can also be used as a second input frequency and this is quite normal now in commercial instruments.

By using harmonics of the local oscillator for use in the mixer, further receive ranges can be arranged. It has to be borne in mind, however, that the conversion loss of the mixer is much greater at harmonic LO inputs. This is the usual practice for commercial analysers which use up to the 20th LO harmonic in order to provide coverage in the millimeter wave range.

The most obvious frequency plan for a receive range of 0 to 1500 MHz would be to locate the IF at above 1500 MHz, e.g. 2 GHz, and the oscillator tuning from 2.0 to 3.5 GHz as shown in **fig. 1.1**. The received input range is then fully covered and the image frequency can easily be eliminated with a simple low-pass input filter.

The snag with this solution is that a tuned oscillator range of 2.0 to 3.5 GHz would be required, which with amateur means, is hardly feasible, moreover, a mixer for these frequencies is relatively expensive.

The IF at 2 GHz presents no real problem but it should be borne in mind that a second conversion to a more amenable frequency for bandwidth and signal processing e.g. 10.7 MHz or 21.4 MHz will have to be carried out. Thus the receiver input range can remain unbroken with a high image rejection and the signal processing can be effectively carried out, all by means of the double-superheterodyne technique. If an IF of 500 MHz is chosen, two ranges of input frequency become available, 0 to 500 MHz and 1000 to 1500 MHz with the local oscillator tuning between 500 and 1000 MHz. The receive ranges may be separated with selectable high and low-pass filters. Diode

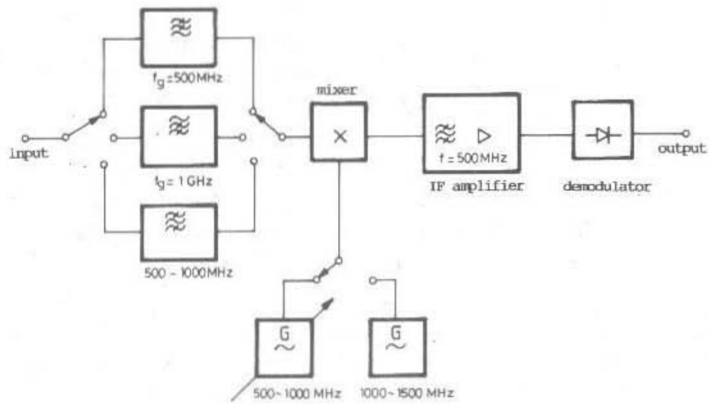

Fig. 1.2: A spectrum analyser tunable in 500 MHz bands

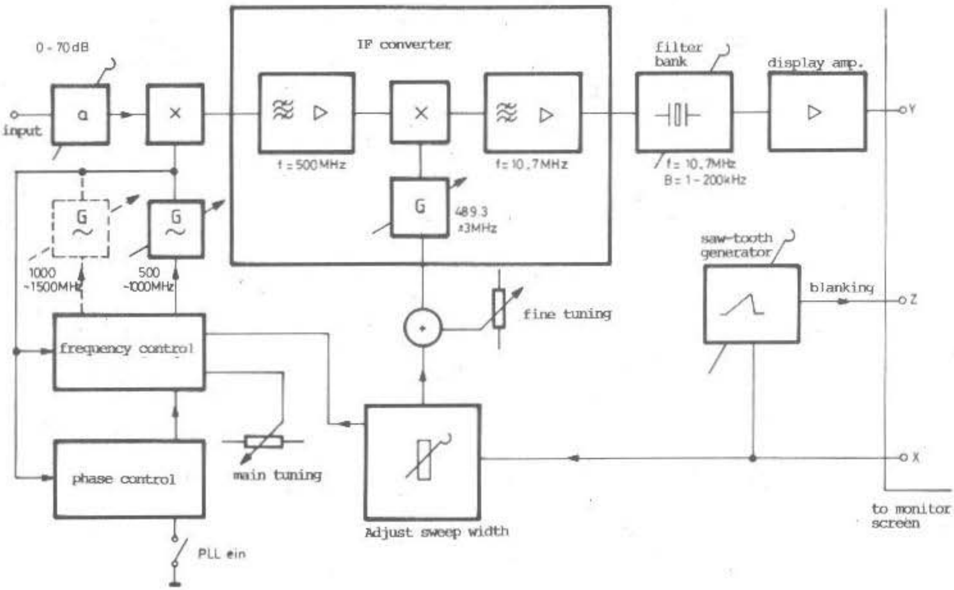

Fig. 1.3: Block diagram of a practical spectrum analyser

tuned VCO's are easily realized at these frequencies and the translation of 500 to 10.7 MHz may be carried out in a single step. The disadvantage of the missing input spectrum range of 500 to 1000 MHz may be overcome by employing an additional 1st LO covering 1000 to 1500 MHz. This further extends the received input range possibility from 1500 to 2000 MHz. The block diagram for this arrangement is shown in **fig. 1.2.**

A third possibility is represented by an IF of 1 GHz but this brings a rather more unfavourable receive range than the above cases as can easily be appreciated.

After settling, in general, that a frequency plan such as that of **fig. 1.2.** is in fact tenable, the block diagram of the spectrum analyser can be fleshed-out a little. This is shown in **fig. 1.3.** which includes the following modules:

- The input mixer: A proprietary mixer such as the SRA-220 may be employed here.
- The voltage-controlled-oscillator, (VCO) controlled by a PLL circuit for adequate frequency stability when using the higher resolutions.
- The IF 2nd mixer down-converting from 500 MHz to 10.7 MHz or 21.4 MHz.
- The main analyser filtering in the 2nd IF.
- The logarithmic display amplifier.
- The control circuits for the tuning and control of the oscillator.

The 2nd mixer, which brings the first IF down to a standard 10.7 MHz or 21.4 MHz, will be considered first.

2. THE 2ND MIXER

Following the plan as laid out in the previous chapter, the 500 MHz 1st IF signals are produced by received inputs being mixed with a first local

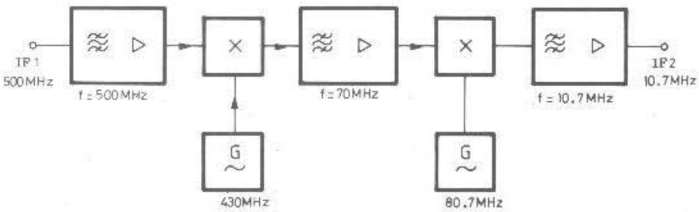

Fig. 2.1: An IF converter with two conversions

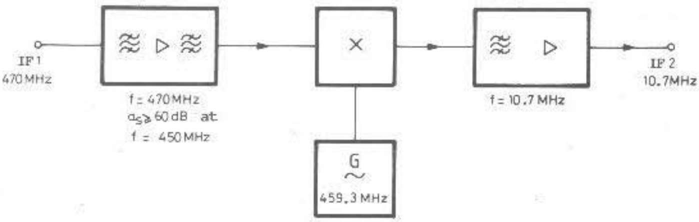

Fig. 2.2: An IF converter with a single conversion

oscillator with a tunable range of 500 MHz to 1000 MHz. The conversion of the signals to 10.7 MHz must be carried out with an image frequency suppresssion of at least 60 dB in order to reduce spurious inputs to the following circuits. In a normal receiver, this would be accomplished using a further IF at about 70 MHz in order to reduce the demands upon the mixer input filter for this level of image suppression. This scheme is shown in **fig. 2.1.**

The disadvantage of this technique is the greatly increased possibility of producing spurious signals from the total of three instrument LOs which would have to be employed. The author has therefore decided upon the direct conversion of 1st IF signals at 500 MHz to an IF of 10.7 MHz and this is outlined in the scheme of **fig. 2.2.** It will be observed that the image frequency is at 21.4 MHz removed from the 500 MHz input frequency and the filter must satisfy the 60 dB image suppression requirement.

This order of suppression can only be achieved with pot resonators or with helix resonators – the latter, on account of its smaller form, is to be preferred. The mini helix filter 10H3 can be obtained from Telequarz, for about DM 20,-. This is tunable from 440 MHz to 500 MHz and at 450 MHz it has a selectivity of some 40 dB. Cascading two of these filters ensures that the image specification of 60 dB suppression will easily be achieved. This entails fixing the first IF not at exactly 500 MHz but at 460 to 470 MHz in order to avoid direct interference from any local channel 21 to 25 television transmitter.

The satellite television receivers use a surface wave filter which has a suitable mid-frequency of 479.5 MHz but the insertion loss at 20 dB is far too high. Also its bandwidth is around 35 MHz with a selectivity of 50 dB. The latter specification would make two in cascade necessary. The author has therefore decided for the helical filter. The detailed schematic of **fig. 2.3** shows two such filters separated by an amplifying BFT 66 transistor. The net gain of this combination is 5 dB with a bandwidth of 6 MHz.

A ring mixer follows the second helical filter which supplies the second IF at 10.7 MHz. The 2nd local oscillator consists of a BF 247 Colpitts oscillator circuit working at 10.7 MHz below the first IF and supplying 10 mW output power. The os-

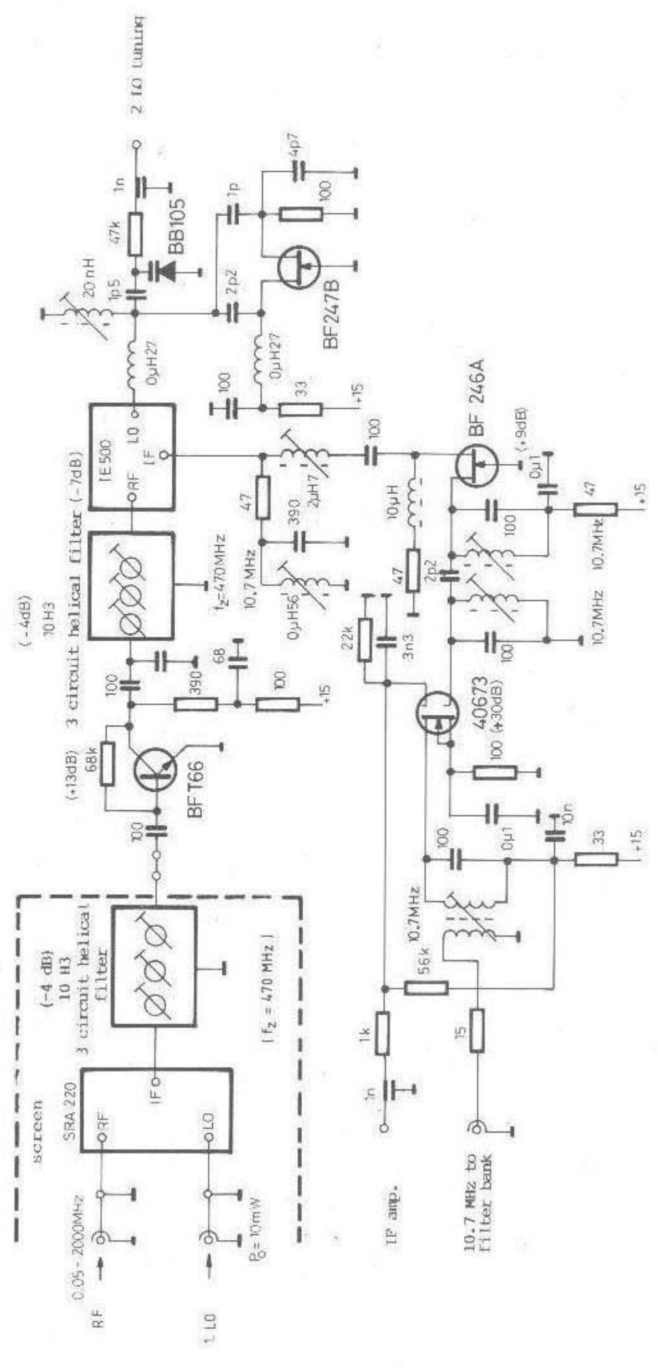

Fig. 2.3: The detailed circuit of the IF converter from 470 MHz to 10.7 MHz

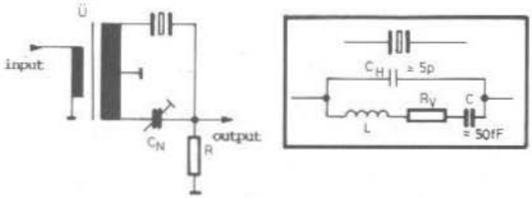

Fig. 3.1:
Basic circuit of a crystal filter

cillator frequency can be shifted a few MHz by means of a varicap diode.

It has been proved that small swept bandwidths, i.e. under 5 MHz, are better accomplished by fixing the 1st LO with a crystal and sweeping the frequency of the 2nd LO. The analyser fine-tuning between the synthesizer step frequencies can also be done here.

The ring-mixer IF output is terminated with a diplexer but it is not strictly necessary. A BF 246 transistor matching stage amplifies the signal by 10 dB. The signal is then filtered by a 10.7 MHz bandpass circuit followed by a further 30 dB of amplification in a MOSFET (e.g. 40673) amplifying stage. The gain of the converter can be controlled downwards from 30 dB by the application of a bias on the gate 2 of this MOSFET stage. The 10.7 MHz amplifying stage has a − 3 dB bandwidth of 500 kHz and a − 60 dB bandwidth of 5 MHz.

For the display of larger bandwidths, 100 to 500 MHz used for example in the investigation of harmonic and spurious signals, this resolution is optimal. The output signal of the IF converter can then be directly fed to the logarithmic display amplifier in order to present a dB-linear display. If a higher resolution is required then an appropriately dimensioned filter can be included in the signal path. The design of suitable filters will now be considered.

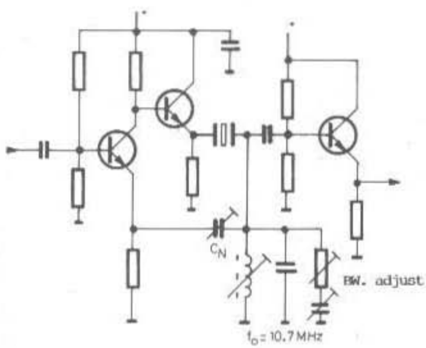

Fig. 3.2:
Crystal filter stage with variable bandwidth

3. THE FILTER BANK

The filter quality determines the spectrum analyser's resolution, i.e. the ability to separate two signals in juxtaposition. Whilst at 10.7 MHz, resolutions of 50 kHz may be obtained with LC circuits, only crystal filters can achieve the lower bandwidths. In fact, it may be stated that one way forward is to select suitable crystal filters from the extensive range offered by firms such as Tele-Quarz or KVG and switch them with relays or diodes into the signal path.

Apart from the fact that each filter costs around DM 150,- they are intended for communication receivers and as such are only of limited use for a spectrum analyser. This is because their steep-sided flanks cause the signal-under-examination to ring as it is swept through the filter. A more suitable passband is bell-shaped, possessing an exponentially falling response equally disposed about the centre frequency in the manner of the so-called Gaus filters. The realisation of such filters is really not so difficult when one is acquainted with the basic circuits of crystal filters. Older amateurs will recall the times when propriety crystal filters were a rarity and filters had to be home-made with surplus crystals.

The basic circuit of a quartz filter is shown in **fig. 3.1.** The input signal is fed in antiphase, from a transformer to a crystal and to a preset capacitor thus forming a bridge circuit. If it is remembered that a crystal is basically a high Q, series-tuned circuit having a parallel holder capacitance then the function of the bridge circuit becomes a little clearer. The capacitive preset arm neutralizes the stray parallel crystal holder capacitance leaving the bandwidth to be determined by the crystal Q and the terminating resistance R. If this preset is made variable, the filter may be tuned over a range of frequencies.

A practical circuit is shown in **fig. 3.2.** The transformer has been replaced by an amplifying stage with two low-impedance outputs and the termination by an LC circuit tuned to the crystal's nominal frequency. This circuit has the advantage that the ever-present unwanted crystal resonances are suppressed completely by this form of termination. An adjustable attenuation, or tuning of the tuned circuit, alters the loading on the crystal and thereby the circuit bandwidth. The output must be loaded by the next stage with a higher impedance.

One such filter stage offers a selectivity of around 20 dB and therefore several must be cascaded. The supply of suitable crystals presents no problems if a quartz filter from a 50 kHz step PLL (from an old NÖBL) transceiver is taken apart and identical crystals extracted. An experimental circuit with four cascaded filters of the type shown in **fig. 3.2.** yielded a 1 to 20 kHz (– 3 dB) tunable bandwidth and a form factor (BW (– 3 dB)/BW (– 60 dB)) of 10 – sufficient for most purposes.

The final version of this filter with four cascaded stages is shown in **fig.3.3.** complete with a by-pass line and final amplification. The bandwidth switching is carried out by means of diodes which switch in the various attenuation networks in parallel with the tuned circuit.

The adjustment for this filter is only possible by the use of a sweep generator test set-up. Most commercial sweepers are of little use for this purpose as they are not able to sweep with precision over such small deviations. The best solution is to make one from a VCO at 10.7 MHz, tuned by a varicap diode over a range of ± 100 kHz. The sweep frequency should also cause the Y amplifier of an oscilloscope to traverse the trace synchronously with it.

Each stage is tuned individually by adjusting the load tuned-circuit for maximum bandwidth. The neutralizing capacitor is then tuned so that the response curve is symmetrical about 10.7 MHz. The fine tuning will be carried out later in the finished analyser for a symmetrical overall response and highest-possible rejection.

The next module in the analyser's signal path to be considered is the logarithmic amplifier.

4. THE LOGARITHMIC DISPLAY AMPLIFIER

It is not normally consequential simply to amplify the IF signal at this stage and to linearly rectify it for presentation to the monitor screen. This would result in signals which have at the most a 20 dB amplitude difference being represented correctly on the same trace. Instead, an amplifier/rectifier arrangement is employed which delivers an output which is proportional to the logarithm of the input voltage. The accuracy of this conversion process determines the level of the instrument.

The actual functioning of this type of chain-rectifier/amplifier will not be considered here as

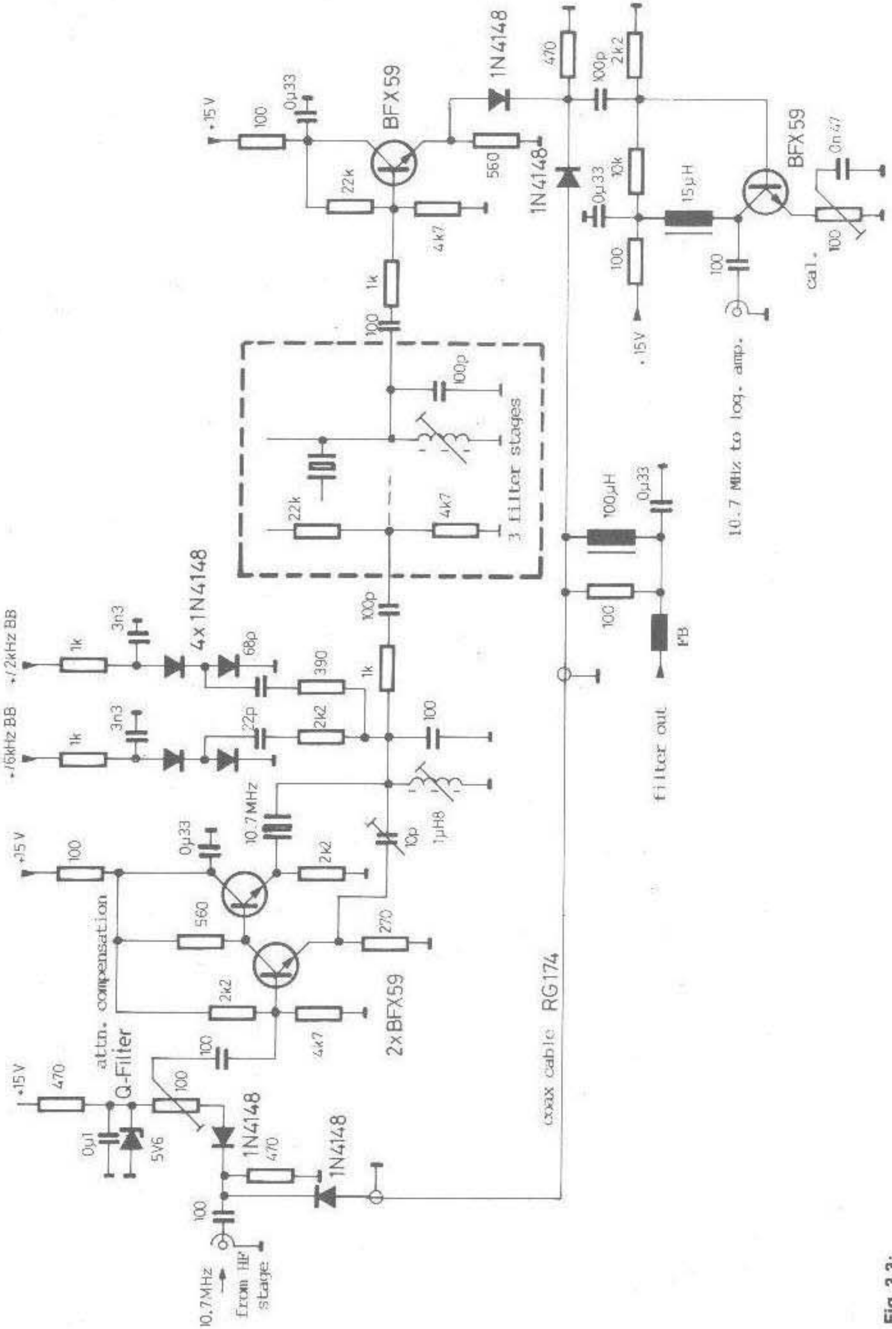

Fig. 3.3:
Spectrum-analyser crystal filter module having four cascaded filters

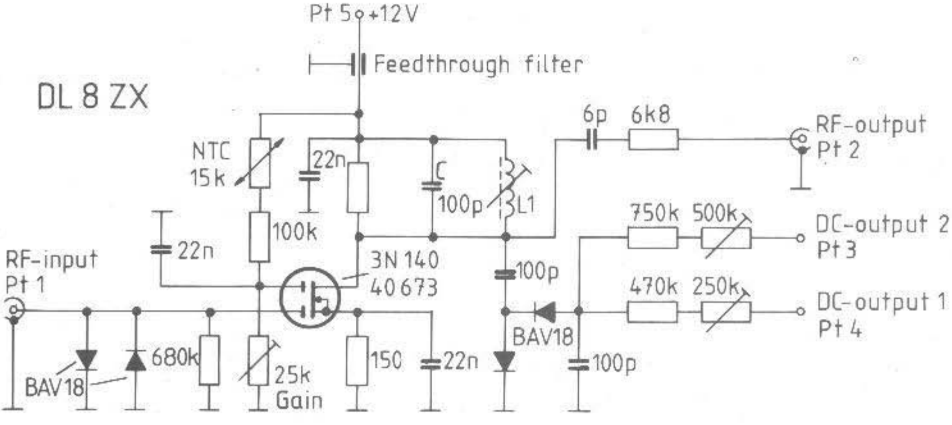

Fig. 4.1: One stage of a chain of rectifier/amplifier (DL 8 ZX)

it has been extensively treated in VHF COMMUNICATIONS 2/1977 by DL 8 ZX. The circuit considered in that article is reproduced in **fig. 4.1**. It is distinguished by its simplicity and easy reproduction. The main disadvantage is that each amplifier has four adjustment points which makes the calibration a bit of a head-ache. Today, semiconductors are much cheaper than they were a decade ago and therefore a trade-in can be made of a little more complexity for ease of calibration. The author obtained his inspiration from an industrial (Hewlett-Packard) logarithmic amplifier and modified it with the following results: –

- The amplifier stages consist of bipolar differential amplifiers whose gain is fixed by resistors and is therefore reproducible and does not need provision for adjustment. No tuning is required as the circuit has no tuned elements so, of course, the selectivity of the display amplifier is minimal.
- Instead of summing the outputs of the instrument rectifiers of each stage the HF voltages of each stage are combined and presented to one instrument rectifier. This obviates any necessity to match individual stage rectifiers. Any deviation from the required demodulator characteristics can be corrected by an adjustment to the summing resistances.
- The linear/dB presentation of the spectrum is effected with only a single rectifier.

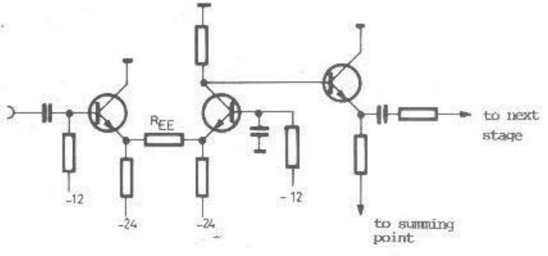

Fig. 4.2: One stage of a logarithmic display amplifier

Fig. 4.2 shows a single differential amplifier stage and **fig. 4.3** the practical circuit having a 70 dB dynamic range. Actually, each differential

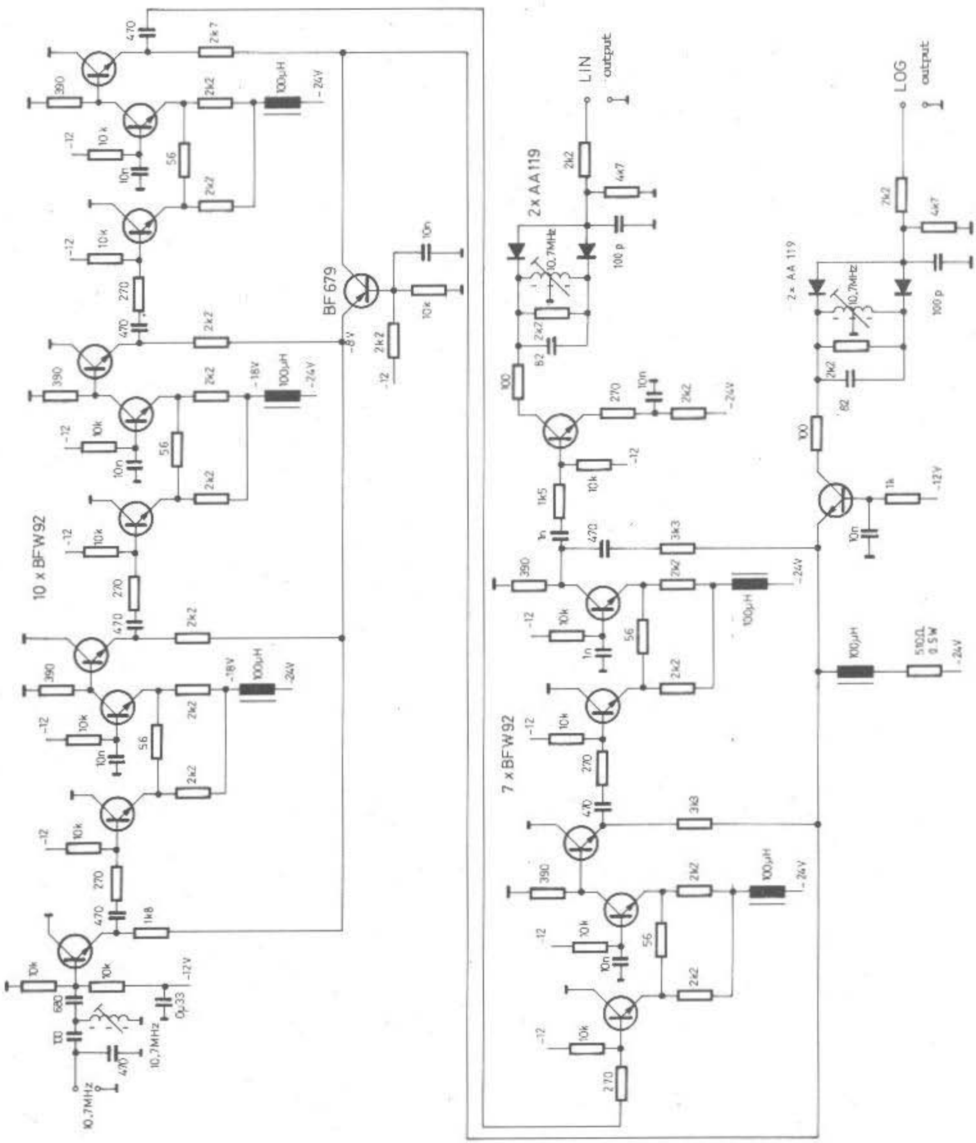

Fig. 4.3:
The logarithmic display amplifier

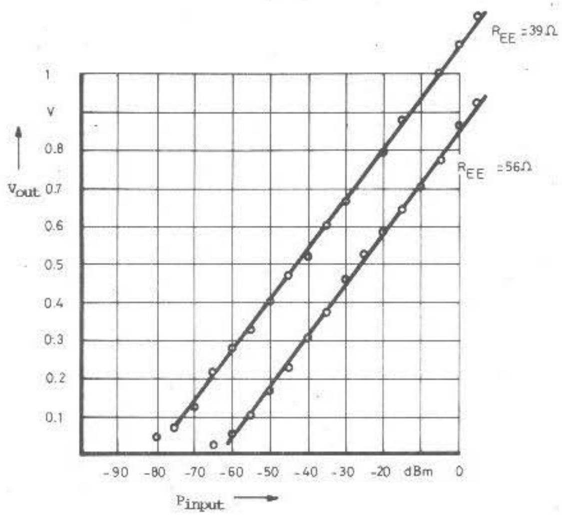

Fig. 4.4: The characteristic curve of display amplifier of fig. 4.3

amplifier stage should have a constant current supply arrangement in its emitter but it has been found that a single high resistance will suffice. The negative rail, however, must be relatively high – in this case – 24 V. Apart from the increased dissipation, this practice has no particular disadvantage.

The level accuracy of such an amplifier lies in the order of 1 to 2 dB absolute, which is considered to

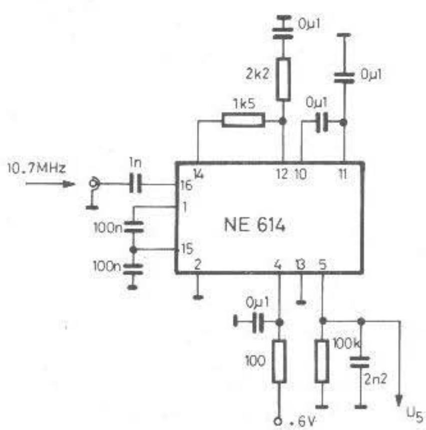

Fig. 4.5a: The NE 614 as a logaritmic display amplifier

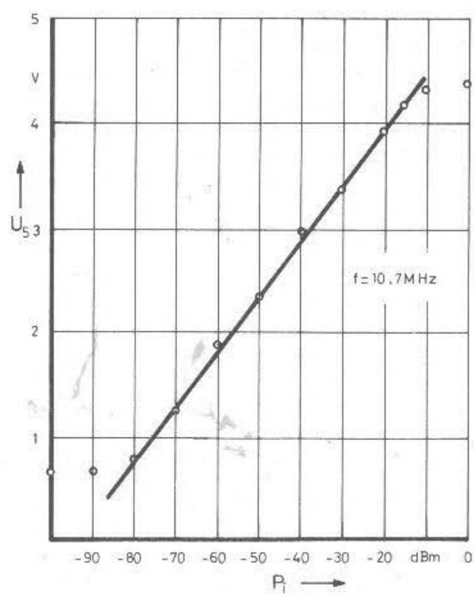

Fig. 4.5b: The characteristic curve of the NE 614 amplifier

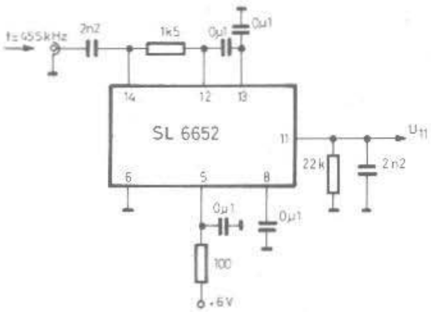

Fig. 4.6a: The SL 6652 as a logarithmic display amplifier

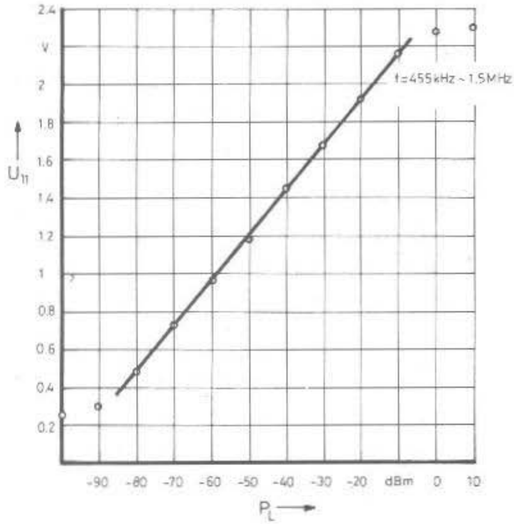

Fig. 4.6b: The characteristic of the SL 6652 log. amp.

be sufficient. The characteristic shown in **fig. 4.4** is that of the prototype amplifier shown in **fig. 4.3.** The accuracy can be improved by decreasing the resistance between each stage and thereby reducing the amplification but there is also a reduction in the dynamic range. If this resistor is dropped from 56 Ω to 39 Ω, for example, over 70 dB of dynamic range can be obtained but the maximum error also increases by ± 1 dB.

A supplementary demodulator, connected to the last amplifying stage, supplies a linear output from an IF input signal. This latter amplifier stage may be connected to the monitor deflection amplifiers and is very useful for more accurate minimal or maximal tuning adjustments as a linear change in level is easier to observe.

It can be seen that both modules can be employed as logarithmic IF amplifier/detectors if the demands upon dynamic range and linearity are not too great.

A further possibility for the realization of a logarithmic IF amplifier is the employment of the purpose-built Plessey differential amplifier series SL 521, SL 523, SL 1521, SL 1523, or SL 1613. All suffer from the disadvantage that they have a large (150 MHz) bandwidth. The wideband noise could be held within bounds by placing tuned circuits between each stage but this would increase the complexity. This circuit technique was not therefore persuad further.

If so much complexity is to be avoided, there are many FM/IF chips which can be used which have a dB-linear characteristic. The VALVO NE 614 is an IF/demodulator chip for frequencies to 15 MHz and the Plessey SL 6652 has an IF limit of only 1.5 MHz but it does have an internal mixer/oscillator circuit at the designer's disposal. The experimental circuit for the NE 614 is shown in **fig. 4.5a** and the IF (10.7 MHz) input signal (dB) versus output voltage is shown in **figs. 4.6a and 4.6b** respectively, but the input frequency in this case was 455 kHz. It can be seen that the SL 6652 has a more linear characteristic but a frequency translation is required from 10.7 MHz to under 1.5 MHz.

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A Spectrum Analyser for the Radio Amateur Part 2 – Concluding

5. THE FIRST LOCAL OSCILLATOR

This oscillator, tuning from 450 MHz to 1000 MHz, has already been described in VHF Communications 4/86 and therefore the details here can be held to a minimum. In this version, shown in **fig. 5.1.**, the oscillator transistor is a BFR 91 and the buffer amplifier an NPN transistor BFG 96, the latter delivering at least 10 mW drive to the following SRA 220 ring mixer. An auxiliary buffer output of 0.5 mW is taken from the twin-hole cored transformer and used for the tuning linearity circuits and for the PLL drive. The components are soldered to the track side of the printed crcuit board, the arrangement being shown in **fig. 5.2.**

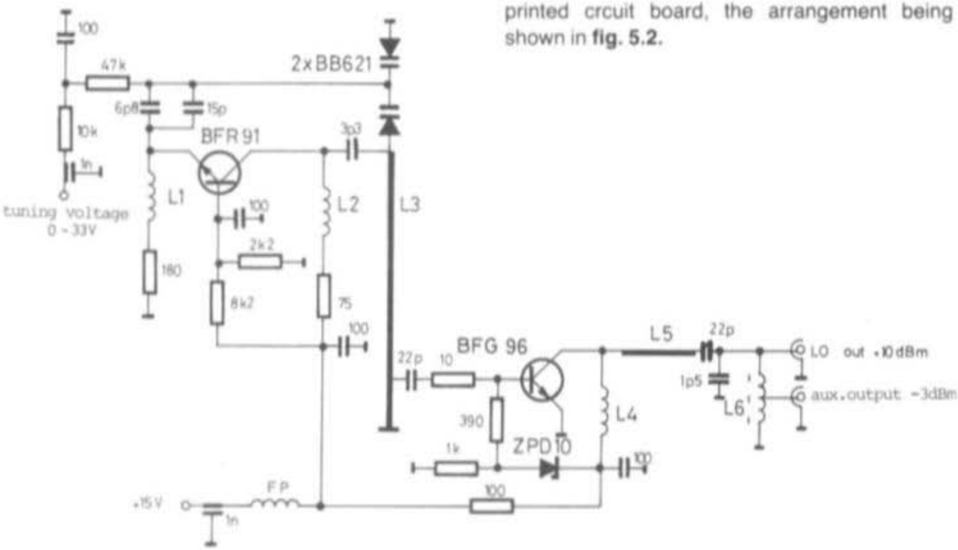

Fig. 5.1.: The first local oscillator having a 450 MHz to 1000 MHz tuning range

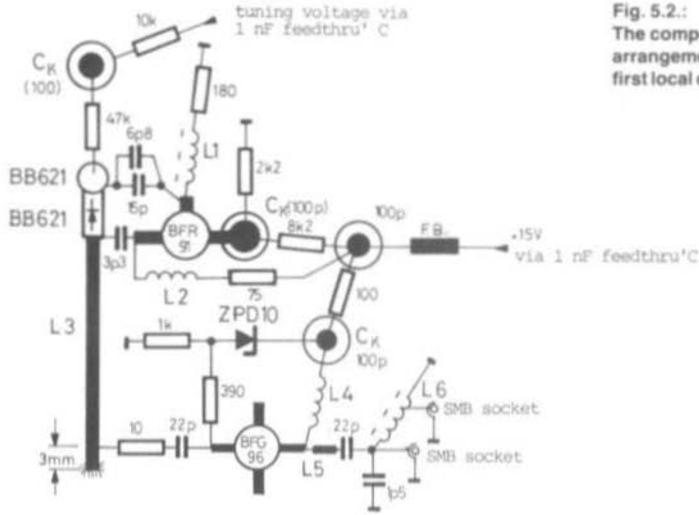

Fig. 5.2.:
The component arrangement of tbe first local oscillator

6. FREQUENCY CONTROL LOOP AND TUNING LINEARITY

The relationship between the tuning voltage and the output frequency of a diode-tuned oscillator normally approximates a square root fuction. In order to obtain a linear relationship between tuning voltage and frequency, it is necessary to employ a resistance-diode network possessing an inverse characteristic, which, when combined with the VCO characteristic, produces the desired linear relationship. The only snag with this arrangement is that each oscillator must be individually adjusted for optimum results.

The employment of a frequency control loop is an alternative method of obtaining the linear relationship between tuning voltage and VCO frequency. A highly linear discriminator is fed from the output of the VCO and delivers information concerning the actual frequency of the VCO. A control amplifier then causes the VCO to return to the set frequency. The disadvantage of this

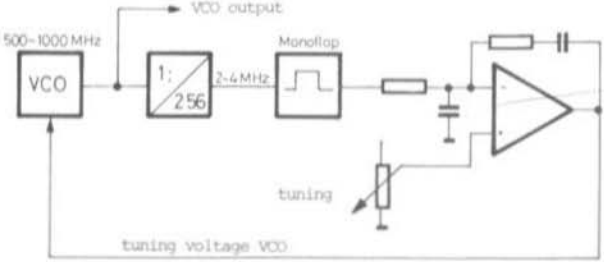

Fig. 6.1.:
Frequency control loop used to linearise the 1st LO's tuning characteristic

method is that the frequency control loop has a finite reaction thus setting a definate time limit to the tuning rate. The diode function generator, on the other hand, suffers no such disadvantage. For the control loop method, a suitable frequency discriminator may be realised using a monostable multivibrator. For every period of input voltage, it delivers an output of constant width and amplitude. The average period of the output pulses determines the output frequency.

As this discriminator can only work up to frequencies of only a few megahertz, the output from the VCO is scaled down with the aid of a simple ECL divider of the type frequently used in commercial TV equipment. The block diagram of the frequency control loop, complete with scaler, is shown in **fig. 6.1.**

The prototype frequency control circuit used by the author is shown in **fig. 6.2.** The VCO main + 10 dBm output goes on to the PLL circuitry and will be dealt with later. The − 3 dBm output from the VCO is fed to a resistive power splitter. It is then buffered by a BFW 92 amplifier before being taken to the SDA 4211 ECL scaler. Following a frequency division by a factor of 256 and a level shift to CMOS, a signal directly derived from the VCO in the range 2 MHz to 4 MHz is available.

The frequency-voltage transducer consists of a monostable 74 HC 123 whose time constants are controlled by a metal-film resistor and an NPO capacitor. The monostable's stability determines the re-setting accuracy and the long-term stability of the frequency control. Since the amplitude of the impulse peaks are determined by the 5 V supply voltage, the high-grade regulator LM 723 is used. This is an order better than the widely used three terminal regulators but the use of a precision regulator such as the REF 02 is not considered to be justified for this application.

The RC network tends to smooth the monostable's output pulses somewhat before they are fed to a low-noise NE 5534 IC amplifier, connected as an integrator or as a PI regulator, which compares the actual with the reference frequency. A transistor amplifier raises the level to that required by the VCO's control voltage

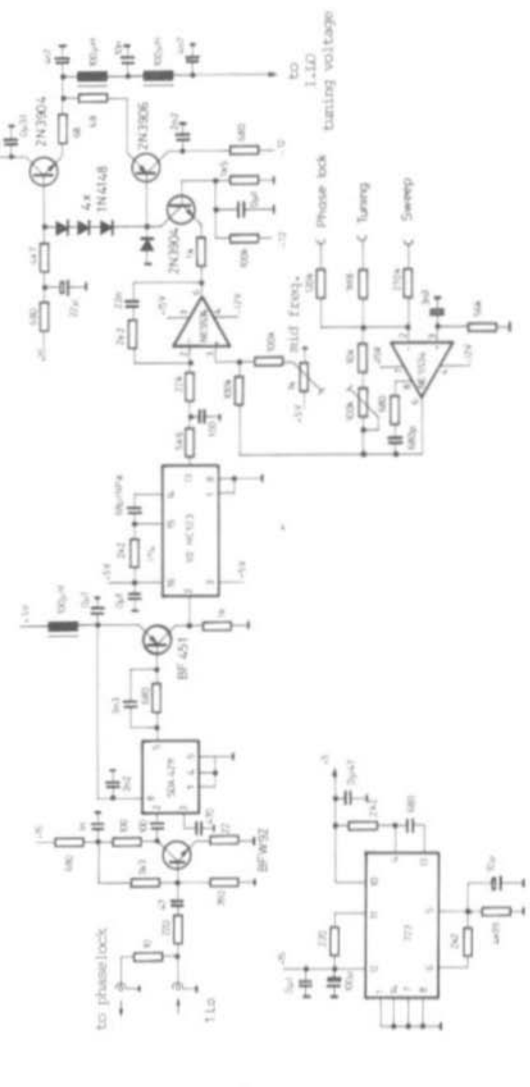

Fig. 6.2.: The first LO's frequency control loop

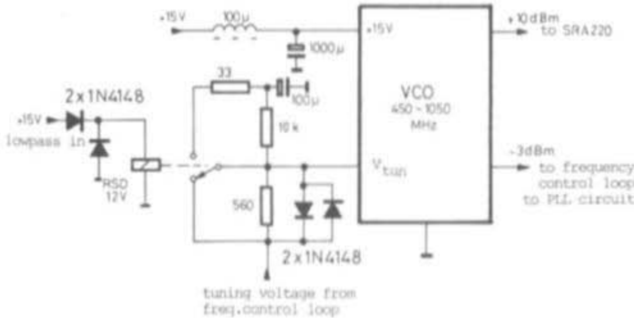

Fig. 6.3.:
The external circuitry of the 1st LO

0 – 30 V and an LC network filters out the residual RF.

Immediately before being applied to the VCO control input, the tuning voltage may be additionally filtered by a relay selected RC network (**fig. 6.3.**). This extra filtering, although improving the noise superimposed on the control voltage, also increases the control loop reaction-time enormously. For this reason it is only activated when the spectrum analyser is being swept at narrow band (200 kHz/div.) from the 2nd local oscillator. The detail is shown in fig. 6.3. The reference frequency for the loop is represented by a weighted sum of the tuning potentiometer voltage, the sweep voltage and the tuning voltage from the PLL circuitry. A second NE 5534 is used here as the summing amplifier. Two multi-turn trimmers adjust the centre frequency and tuning sensitivity.

With the dimensioning as suggested here, the frequency control loop has a 200 Hz bandwidth. This allows a loop reaction time to tuning voltage variations of 1.5 ms. The VCO frequency stability is better than 100 kHz with this control loop and it is, in most cases, largely dependent upon the externally generated tuning voltage.

A static measurement of linearity using a frequency counter and a digital voltmeter reveals practically no errors. The VCO exhibits a residual modulation, i. e. from hum and LF noise, of some 30 kHz peak. Only on the filter bandwidth from 1 to 2 kHz can it be observed that a better VCO stability would be desirable. This theme is persued in the following chapter which introduces a PLL circuit which may be switched to sweep presentations of 200 kHz / div. or less.

7. THE PHASE LOCKED LOOP

When swept displays of under 200 kHz / div. are required, the spectrum analyser presented here sweeps the 2nd LO at 460 MHz whilst the 1st LO is free-running. In order to reduce the effects of the 1st LO's noise modulation, the oscillator is controlled from a crystal frequency reference. The author attempted to use a television receiver PLL synthesizer for this purpose but it proved unsuitable. Test measurements showed that the noise from a television synthesizer was greater than that from the frequency control loop described above. The trouble lies in the large division factor of 64 or 256 in the TV tuner and its inherently low phase comparator frequency.

The PLL circuit is therefore so dimensioned that the smallest possible VCO division factor is used.

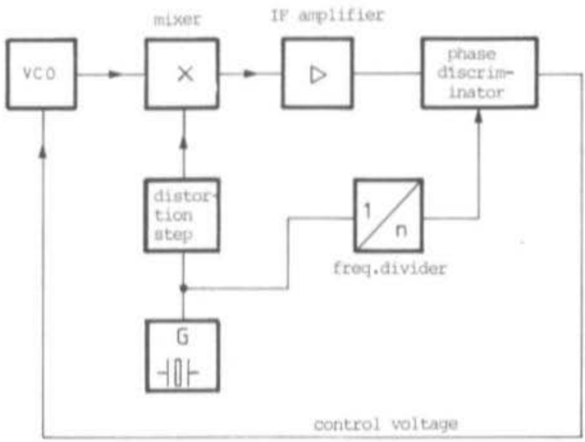

Fig. 7.1: 1st LO PLL block diagram

In the interests of easier operation, a control technique, which allows manual tuning between fixed spot frequencies, is to be preferred. One possibility is a PLL circuit with harmonic mixing as shown in block diagram (**fig. 7.1.**). The VCO signal is mixed with a harmonic-rich crystal oscillator. The resulting intermediate frequency is amplified and taken to a phase / frequency comparator which controls the VCO tuning voltage.

A problem crops up here with the production of the harmonic signals as the spectrum must extend well over the 1 GHz region. The author found that an harmonic spectrum up to about 300 MHz could be achieved relatively easily by using the gate times of TTL circuits. This offers the possibility of dividing the VCO frequency to under 300 MHz and then to arrange the harmonic mixing.

The final PLL circuit is shown in **fig. 7.2.** It may be seen that the input again possesses a resistive divider which passes the greater proportion of the input power to the LO output on the front panel. A smaller portion is amplified by a two-stage transistor (BFR 15) amplifier and taken to the CA 3199 ECL scaler.

The price of integrated UHF circuits has fallen recently (to less than 10 DM) making their use almost mandatory. For a start, the Mini-Circuits MAR-8 is suggested. The input signal is divided by 4 in a reasonable priced (6 DM) CA 3199 frequency scaler which is capable of working up to 1250 MHz. The output from the divider is taken via a low-pass filter to a mixer which is arranged in a sample-hold FET circuit.

A CD 4060 crystal oscillator circuit working at 4 MHz is divided by 16 and presents a 256 kHz square wave into positive going needle pulses about 2 ns wide. The needle pulses gate the FET switch. A further division of the crystal frequency produces a 32 kHz reference for the frequency / phase comparator.

The IF signal is taken at high impedance from the mixer by a FET buffer and then on to the CD 4046 frequency / phase comparator via a 50 kHz low-pass filter. A control voltage from the CD 4046 pin 13, is eventually taken to the PLL after being filtered and buffered. A BF 246 FET switch passes the PLL control voltage on to the phase-lock input of the frequency control loop (**fig. 6.2.**). The 500 Ω preset is adjusted to optimise the PLL sensitivity. A lock-detect signal is also available from the CD 4046 pin 1. This is made to drive an LED located on the front panel, marked "LOCK", and signals the correct functioning of the PLL.

When initially switched on, it was found that the PLL did not lock at or near the selected spot fre-

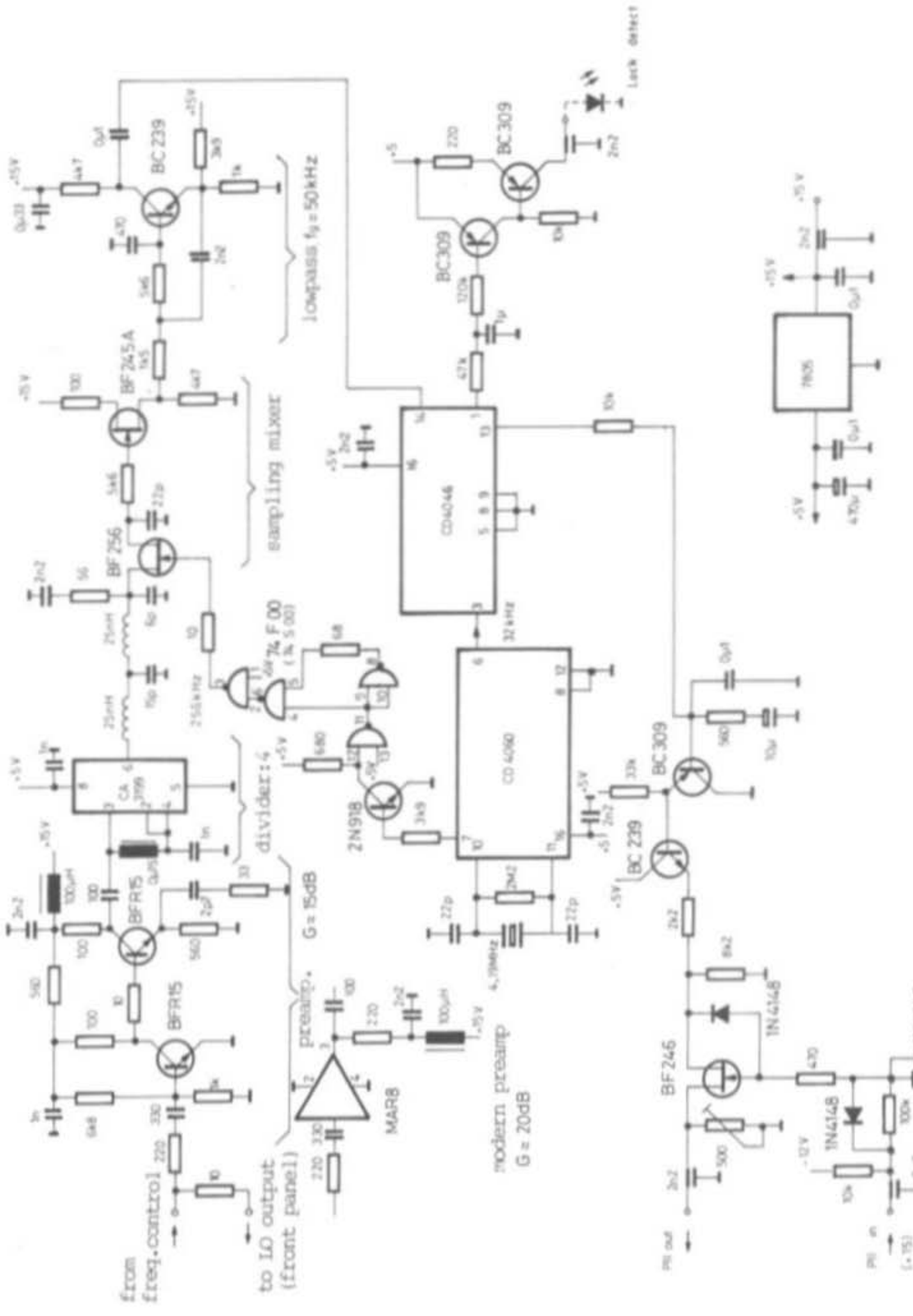

Fig. 7.2.: Detailed circuit diagram of 1st LO PLL

quency but locked at some other random frequency far removed. This error was corrected by arranging the PLL to turn on gradually upon switch-on thus avoiding switching transients. This was accomplished by including an RC network in the gate circuit of the BF 246 switching FET. This has the effect of gradually raising the gate voltage upon switch-on and thereby the PLL loop gain.

Using the circuit proposed, here the frequency-comb, spot-frequencies are spaced by 1 MHz. This presents no problems for the fine tuning between spot frequencies; this being done at 5 MHz by a second oscillator. The frequency stability of the spectrum analyser is only determined by the 1st LO's reference crystal stability and the drift from the fine-tune, second oscillator. When thermally stabilized, the frequency drift of the analyser is less than 10 kHz. In operation, the analyser is first switched to a sweep display of 200 kHz per division, at about the frequency of interest, and then the PLL is switched on. By careful correction of the main tuning, the PLL is brought into lock as indicated by the illumination of the "lock" LED. Now, the desired signal can be fine-tuned and the sweep width further reduced for a narrow-band display. The main tuning should not then be altered otherwise the 1st LO will lock at a neighbouring frequency and the display lost. If a display of 500 kHz per division, or greater, is required the PLL is switched off automatically.

8. OTHER MODULES

A few other circuit details must be mentioned in order to complete the description of a practical spectrum analyser. Some modules e. g. the power supply delivering + 15 V, + 35 V, − 12 V, − 24 V, the saw-tooth generator for the sweep voltage and deflection amplifiers have been deliberately omitted as these items are heavily dependent upon the type of display tube which has been utilised.

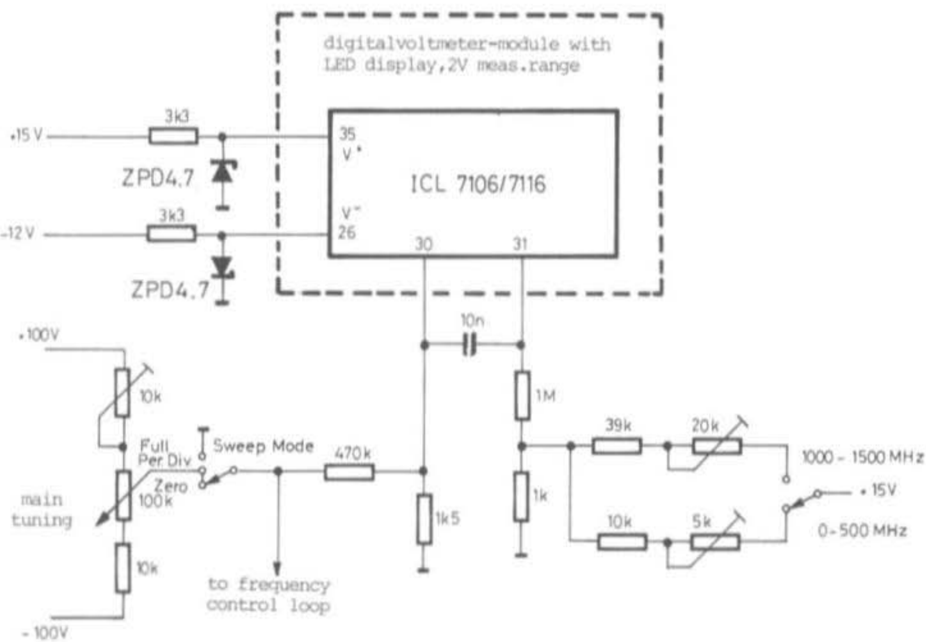

Fig. 8.1.: Arrangement of the tuning voltage and frequency display

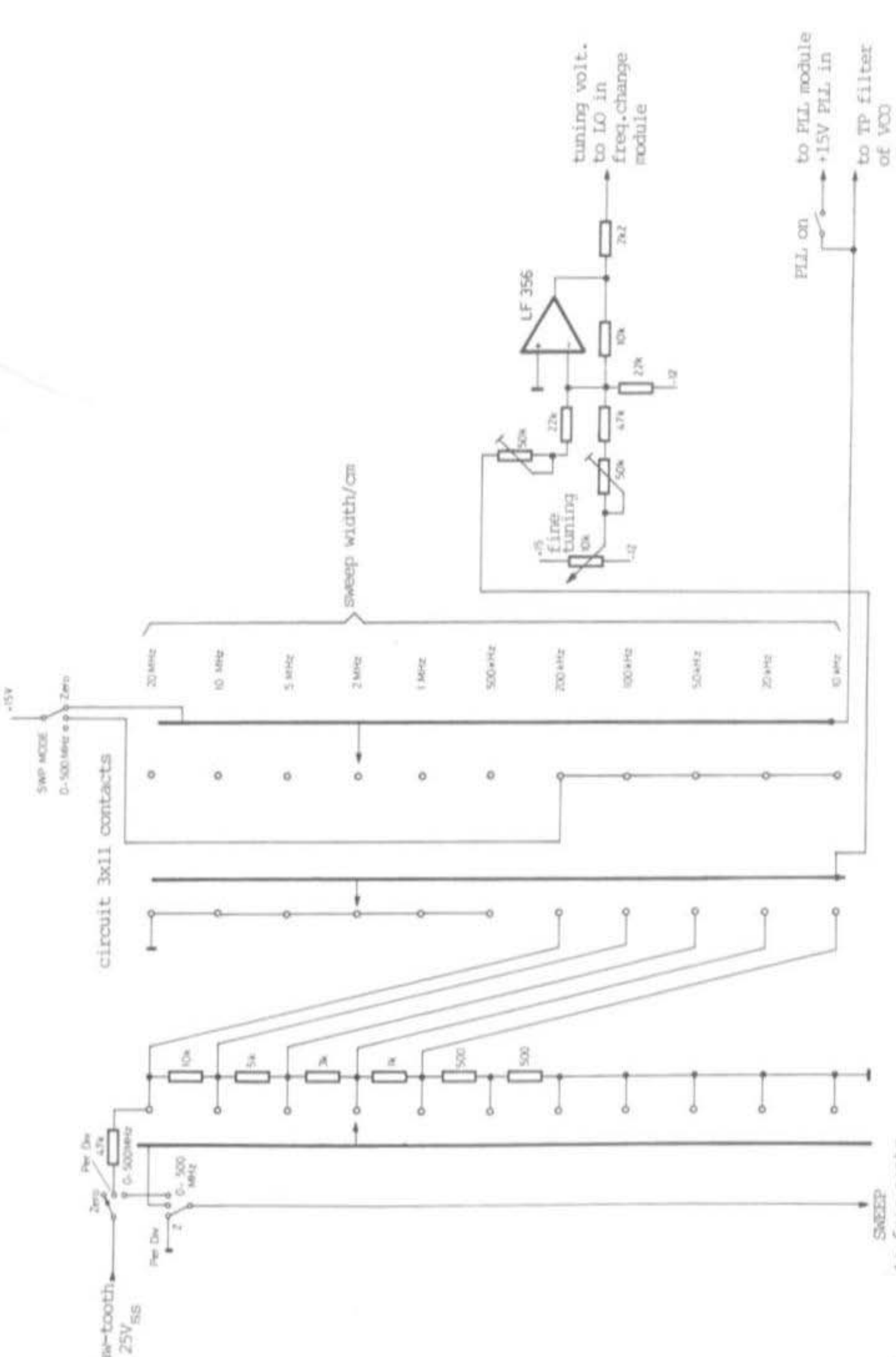

Fig. 8.2.: The switching logic for the tuning voltages of the 1st and 2nd LOs

The circuit details to be considered now are the production of the tuning-voltage supply, the digital frequency indicator and the support module for the display sweep-width. A 100 kΩ, 10-turn potentiometer serves as the tuning potentiometer. The tuning voltage, in the author's version was drawn from a highly stable ± 100 V supply which was available in the instrument. This voltage range is not strictly necessary as the tuning range can be altered by suitable dimensioning of the summing resistors in the frequency control loop circuit to suit the available tuning voltage supply to the tuning potentiometer. The main thing to watch, is that the supply itself is absolutely stable and free of hum and noise.

A digital voltmeter can be used as the frequency display which is actually measuring the tuning voltage. The author used a 3 1/2 digit voltmeter with an LCD display and a 2 volt range. This used the well-known CMOS module ICL 7106 or ICL 7116 by GE-Intersil. This unit is equipped with a floating differential indicator and has a consumption of only 2 mA! As opposed to the usual 9 V battery supply, a mains derived source is used in a full-wave circuit to two Z-diodes delivering ± 4.7 V w.r.t. ground.

When constructing this item, or using a modified commercial digital voltmeter, two points should be considered. If the module originally used 9 V batteries then the minus (pin 30) is strapped to ground (pin 32). This connection is to be interrupted.

This IC sometimes exhibits problems with common-mode voltages which drive the integrator into a premature run-down condition. This has the effect of the display indicating up to a certain input voltage and then saturates, or sticks, at that indication. This effect can be countered by increasing the value of the integration resistance at pin 28 or increasing the value of the integration capacitor at pin 27.

Figure 8.1. shows the external circuitry used by the author to support the ICL 7116 in the area of the signal input. By means of a selector switch, either one of two receive input ranges may be selected 0 to 500 MHz or 1000 to 1500 MHz and correctly indicated. Operating this switch can also be caused to switch in the appropriate high and low-pass filters before the first mixer, but the author did'nt do it in this manner in the prototype instrument. The frequency indication has an accuracy of a few MHz which is sufficient for most purposes. If a higher accuracy is desired, it can be calibrated with a spectrum generator.

The sweep voltage, employed to wobbulate the oscillators of the spectrum analyser, is a symmetrical about zero, saw-toothed wave-form of 25 V peak to peak. The frequency of the wave-form is variable between 2 Hz and 25 Hz. The generation of the various scan widths is given in **fig. 8.2.** By means of a stepped divider sweep widths of 10 kHz/div. and 20 MHz/div., in a 1-2-5 sequence, may be selected. For scan widths over 500 kHz/div., the 1st oscillator is deviated with an attenuated portion of the sweep voltage applied to the frequency control loop circuit. When smaller scan widths are selected, only the second oscillator is wobbulated. Also, the sweep-voltage and the voltage from the fine-tune potentiometer are summed in an operational amplifier and taken to the 2nd LO's varicap diodes in the IF converter module. A correction of the tuning characteristic was found to be unnecessary. In this mode of operation, the control speed of the frequency regulator loop is reduced in order to attenuate the 1st LO's noise modulation. In addition, the PLL stabilization can be switched in.

By means of a further switch a display over the whole 500 MHz range can be selected – the mid-frequency being fixed. Another facility is that the whole wobbulation system can be switched off in order to observe the modulation on a single signal.

9. CONSTRUCTIONAL HINTS

There are many and various ways by which the described modules may be connected together to form a complete spectrum analyser. The most elaborate way is to make the display and power supplies in order to build a self-contained instrument. Life is also made easier when a ready-to-hand oscilloscope can be pressed into service

for the display. The oscilloscope's time-base can then be used for the spectrum analyser's saw-tooth waveform requirement. Most oscilloscopes have an external connector for the time-base voltage but if not, one can be easily fitted. Also, with any luck, the oscilloscope's power supplies will also supply the analyser's power requirements as well.

The owners of an oscilloscope having a plug-in unit capability are even better off. One of the plug-in units can be stripped and the chassis used to build the spectrum analyser circuits. The oscilloscope can then be used for its original purpose, or for the spectrum analyser facility, as desired. The author chose this method using a Hewlett Packard 140 A oscilloscope main frame and one of the little used plug-in options being turned into an analyser plug-in. The main frame could deliver + 250 V, + 100 V, − 12 V and − 100 V. The + 35 V and − 24 V supplies required by the analyser were easily derived from the +/− 100 V oscilloscope supply, using resistors and zenner diodes.

The + 15 V supply had, however, to be a little more elaborate because of the 500 mA consumption. One possibility, offered by all older plug-in oscilloscopes, is the employment of the 6.3 VAC heater voltage which is brought out at the main plug for the plug-in unit. Using a small laminated transformer, a full-wave rectifier and a three-terminal regulator, the necessary voltage may be made available.

The HP 140 does not have a deflection amplifier, i.e. the deflection card of the CRT lies directly at the plug-in's plug. Two deflection amplifiers are necessary therefore but the bandwidth need only be about 50 kHz. The author used for this purpose a ready-to-hand board which also contained a saw-tooth generator and a blanking amplifier which suppressed the trace return.

It is not necessary to go into further details as they are largely dependent upon the type of display device used by the author. Other equipment must use an individual method for utilising it to combine with a home-made spectrum analyser. Before attempting to modify an oscilloscope for this purpose, it is recommended that the oscilloscope handbook is given a thorough persual. Details on deflection amplifiers may be found in

Fig. 9.1.: The author's spectrum analyser

volume 4 of the series "Professional Electronics" published by Francis Verlag.
The author's prototype instrument is shown in **fig. 9.1.**

10. ADDITIONS

The spectrum analyser as described may be given a few further facilities. A few suggestions are given as follows:

– An additional VCO tuning from 1000 MHz to 1500 MHz will enable the input range to be extended from 500 MHz to 1000 MHz. The ECL divider, in the frequency control loop and in the PLL, are well able to work at these frequencies as tests have shown.

– A frequency converter or an external harmonic mixer for the microwave bands can be connected.

– A sweep generator having a 60 dB dynamic range and a tuning range of 0 – 500 MHz may be made by synchronising another oscillator so that it is swept in step with the spectrum analyser's 1st LO.

– Using high and low-pass filters, automatically switched with the input range switch, reduces the possibility of receiving spurious signals in the display trace.

– The use of a better, i.e. noise reduced, local oscillator will improve the instrument's dynamic range. This is extremely important when observing a number of closely arrayed signals with large level differences between them.

It may be seen that the spectrum analyser offers many possibilities for individual expansion schemes and for experimentation for improved performance with the basic circuits. The purpose of this article, as stated at the start, was to generate the necessary interest to begin a project of this kind and also to dispel the unfounded fears attending its complexity.